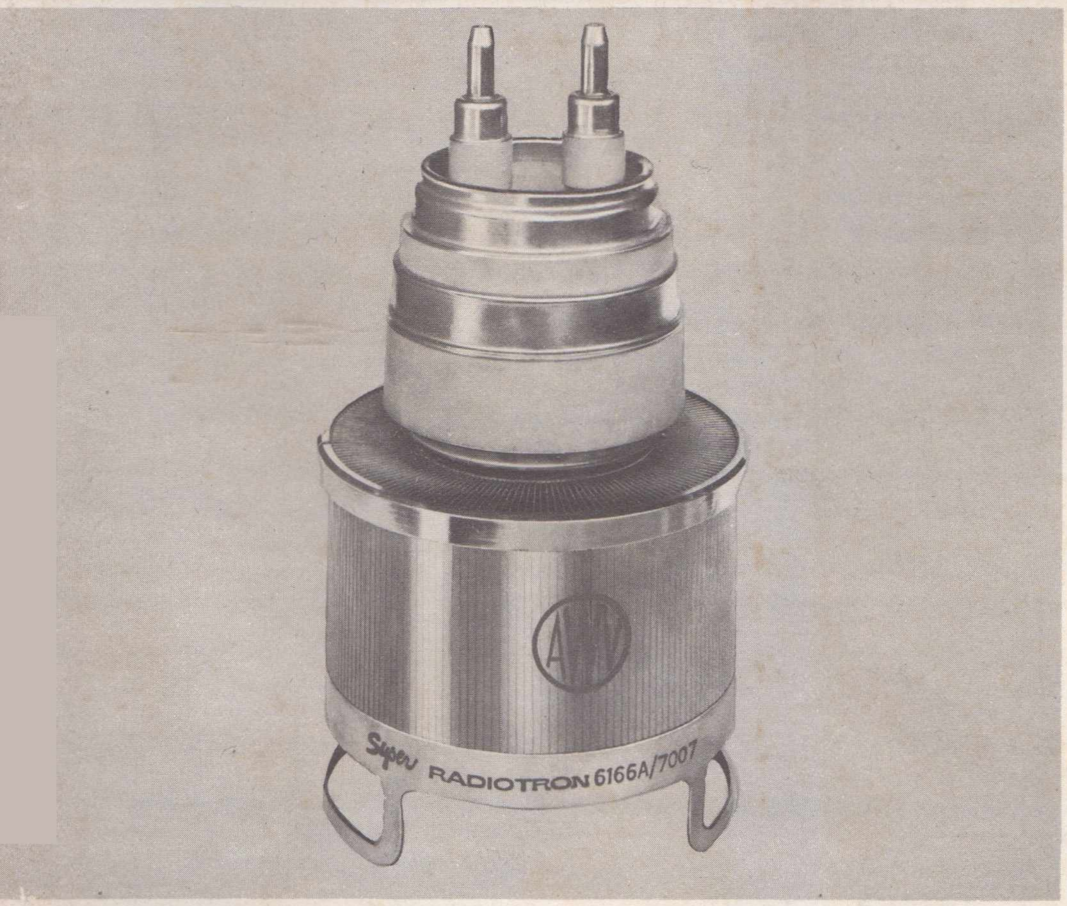
RADIOTRONICS



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COVER:

AWV Super Radiotron 6166A ceramic beam power tetrode.

Vol. 32, No. 4 November, 1967

4

AN AWV PUBLICATION

REGISTERED IN AUSTRALIA FOR TRANSMISSION BY POST AS A PERIODICAL

AS300, AS301, AS302

AWV SILICON N-P-N PLANAR EPITAXIAL TRANSISTORS

The AS300, AS301 and AS302 are n-p-n silicon planar epitaxial transistors in an epoxy package. They are designed for general high frequency applications e.g. amplifiers mixers or oscillators.

Absolute Maximum Ratings

	AS300	AS301	AS302	
Collector-base Voltage	25	25	25	V
Collector-emitter Voltage (Base emitter resistance = 100 KΩ)	15	25	25	V
Emitter-base Voltage	3.5	3.5	3.5	V
Emitter current	50	50	50	mA
Base current	25	25	25	mA

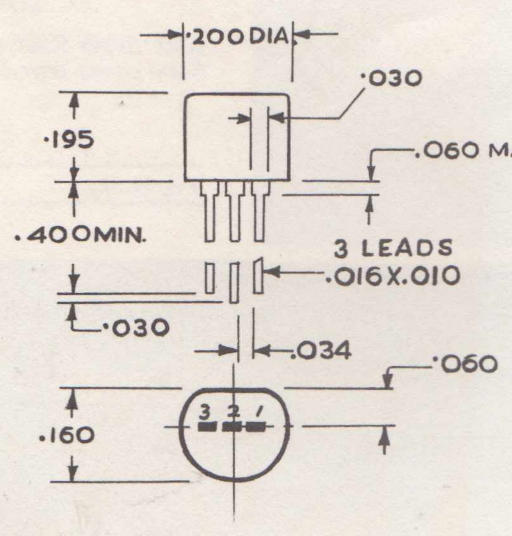
Thermal Ratings AS300, AS301, AS302

Dissipation in an ambient temperature up to 25°C .... 200mW max.
Derate linearly to zero at 135° C.

During soldering lead temperature must not exceed 255°C. for 10 seconds maximum within 1/16" of can.

Storage temperature – 25°C to 135°C.

DIMENSIONAL OUTLINE

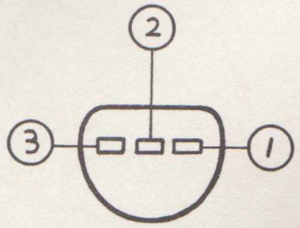


Dimensions in inches

TERMINAL DIAGRAM

Dimensions in inches

TERMINAL DIAGRAM



Lead 1-Emitter
Lead 2-Base
Lead 3-Collector

Electrical Characteristics at TA = 25°C.

CHARACTERISTICS	SYMBOL	TEST CONDITIONS	AS300			AS301			AS302			UNITS
			MIN.	TYP.	MAX.	MIN.	TYP.	MAX.	MIN.	TYP.	MAX.	
COLLECTOR CUT OFF CURRENT	I_{CBO}	$V_{CB} = 1$ V	–	–	20		–	20	–	–	20	nA
		$V_{CB} = 25$ V	–	–	1	–	–	1	–	–	1	μA
EMITTER CUT OFF CURRENT	I_{EBO}	$V_{CB} = 3.5$ V	–	–	1	–	–	1	–	–	1	μA
COLLECTOR-EMITTER VOLTAGE	V_{CER}	$I_C = 100\ \mu$A, $R_{BE} = 100$ KΩ	15	–	–	25	–	–	25	–	–	V
STATIC FORWARD CURRENT TRANSFER RATIO	h_{FE}	$V_{CE} = 6$ V, $I_E = 6$ mA	27	–	270	27	–	90	70	–	270	–
GAIN BANDWIDTH PRODUCT	f_T	$V_{CE} = 6$ V, $I_E = 2$ mA	–	750		–	750	–	–	750	–	MHz
COLLECTOR TO BASE FEEDBACK CAPACITANCE	C_{cb}	$V_{CB} = 8$ V, $I_E = 0$	–	–	1.5	–	–	1.4	–	–	1.4	pF

AS310, AS311, AS312, AS313

AWV SILICON N-P-N PLANAR EPITAXIAL TRANSISTORS

The AS310, AS311, AS312 and AS313 are n-p-n silicon planar epitaxial transistors in an epoxy package for audio frequency applications. The AS310 and AS311 feature low saturation voltage (less than 0.3 volts at I_C = 300 mA).

Absolute Maximum Ratings

	AS310	AS311	AS312	AS313	
Collector-base voltage	40	20	40	20	V
Collector-emitter voltage (Base open)	40	20	40	20	V
Emitter-base voltage	5	5	5	5	V
Emitter current	350	350	350	350	mA
Base current	30	30	30	30	mA

Thermal Ratings AS310, AS311, AS312, AS313

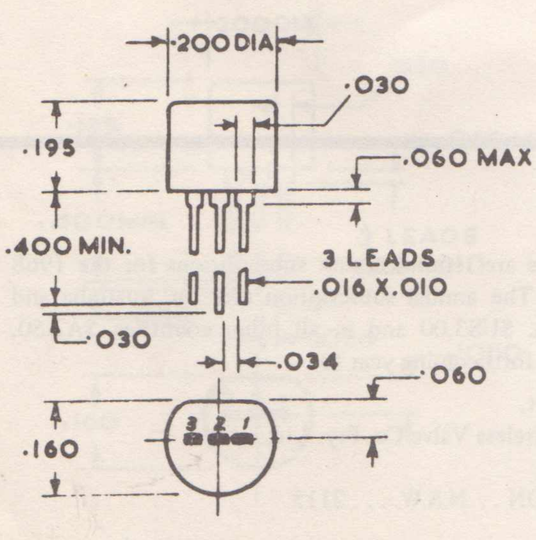
Dissipation at case temperature up to 25° C .... 700mW max.
Derate linearly to zero at 135° C.

Dissipation in ambient temperatures up to 25° C .... 300mW max.
Derate linearly to zero at 135° C.

During soldering lead temperature must not exceed 255°C for 10 seconds maximum within 1/16 of case.

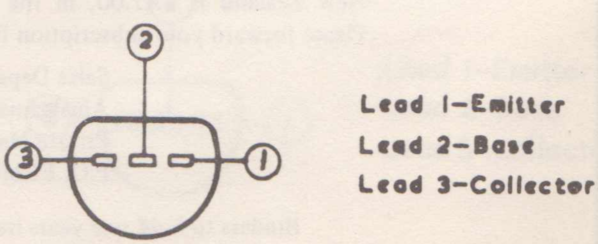
Storage temperature range −25°C to 135° C.

DIMENSIONAL OUTLINE



Dimensions in Inches

TERMINAL DIAGRAM



AS310, AS311, AS312, AS313 (cont'd)

Electrical Characteristics at TA = 25^{o}C.

CHARACTERISTICS	SYMBOL	TEST CONDITIONS	AS310			AS311			AS312			AS313			UNITS
			MIN.	TYP.	MAX.	MIN.	TYP.	MAX.	MIN.	TYP.	MAX.	MIN.	TYP.	MAX.	
COLLECTOR CUTOFF CURRENT	I_{CBO}	V_{CB} = IV	–	–	20	–	–	20	–	–	20	–	–	20	nA
		V_{CB} = 12V	–	–	–	–	–	500	–	–	–	–	–	500	
		V_{CB} = 25	–	–	100	–	–	–	–	–	100	–	–	–	
EMITTER CUT OFF CURRENT	I_{EBO}	V_{EB} = 2.5 V	–	–	100	–	–	500	–	–	100	–	–	500	nA
EMITTER-BASE VOLTAGE (COLLECTOR OPEN)	V_{EBO}	$I_B = 50 \mu A$	5	–	–	5	–	–	5	–	–	5	–	–	V
COLLECTOR-EMITTER VOLTAGE (BASE OPEN)	V_{CEO}	I_C = 10 mA	40	–	–	20	–	–	40	–	–	20	–	–	V
COLLECTOR-EMITTER SATURATION VOLTAGE	$V_{CE(Sat)}$	I_C = 300 mA, I_B = 15 mA	–	–	300	–	–	300	–	–	–	–	–	–	mV
		I_C = 100 mA, I_B = 5 mA	–	–	–	–	–	–	–	–	200	–	–	200	
BASE-EMITTER SATURATION VOLTAGE	$V_{BE(Sat)}$	I_C = 300 mA, I_B = 15 mA	–	–	1.5	–	–	1.5	–	–	–	–	–	–	V
		I_C = 100 mA, I_B = 5 mA	–	–	–	–	–	–	–	–	1.3	–	–	1.3	
STATIC FORWARD CURRENT TRANSFER RATIO	h_{FE}	V_{CE} = 10 V, I_E = 10 mA	60	–	300	60	–	300	60	–	300	60	–	300	
GAIN BANDWIDTH PRODUCT	f_T	V_{CE} = 10 V, I_E = 10 mA	–	175	–	–	175	–	–	175	–	–	175	–	MHz
COLLECTOR TO BASE FEED BACK CAPACITANCE	C_{cb}	V_{CE} = 6 V, I_E = 0	–	13	20	–	13	20	–	13	20	–	13	20	pF
THERMAL RESISTANCE JUNCTION TO CASE	$R_{\Theta J-C}$		–	–	150	–	–	150	–	–	150	–	–	150	°C/W

NEWS & NEW RELEASES

R.C.A. INTEGRATED CIRCUITS

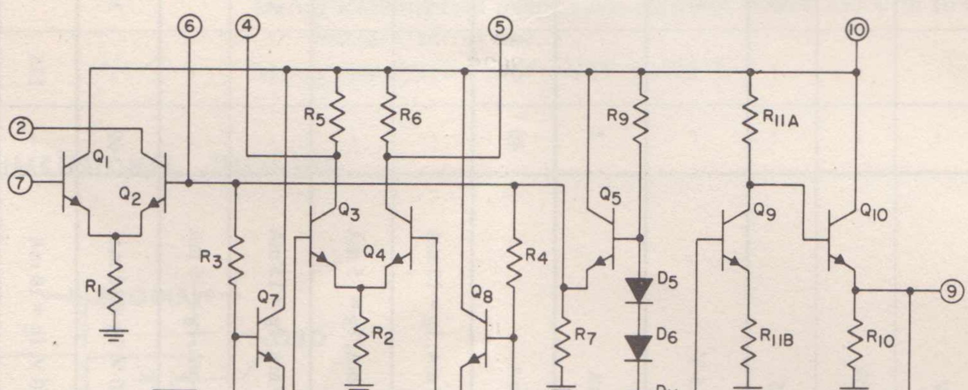
A further three linear integrated circuits have recently been added to the comprehensive range of devices now available.

CA3034

This circuit is a high frequency wide-band amplifier/phase detector. Features of it include:

- Differential input amplifier
- Dual Phase Detector
- Differential output amplifier
- Internal regulated bias supply
- Compensated reference voltage supply

The CA3034 has been characterised specifically for automatic frequency applications in T.V. receivers, FM Tuners and frequency synchronizer systems (e.g. computer clocks) but is suitable for any application where correction voltages are required as a function of frequency.



Schematic Diagram for CA3034

CA3035

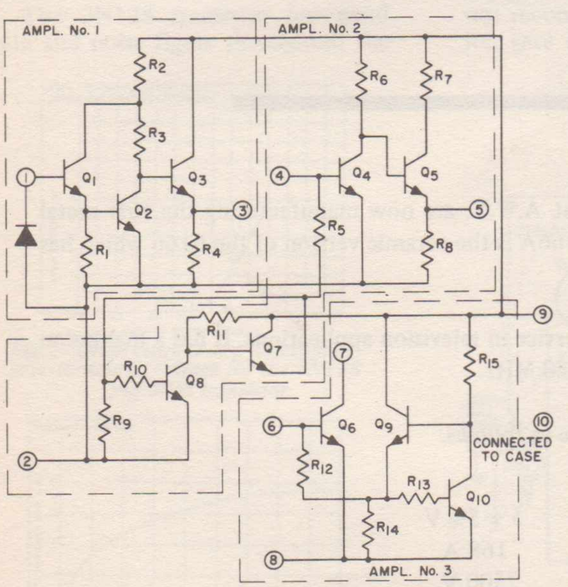
Features of this circuit which is an ultra high gain wide band amplifier array include:

- Three separate amplifiers
  - –can be operated independently or in cascade
  - –gain and bandwidth for each can be adjusted with suitable external circuitry
- Low noise performance
- High frequency response
- Only one power supply required
- Emitter-follower outputs on first two amplifiers

  Un-committed collector on last amplifier for a maximum external load flexibility.

There are many applications for which the CA3035 is suitable as its three amplif- can be used individually or in cascade. Some examples of its application are:

Remote control amplifiers
Audio pre-amplifiers
AC coupled IF amplifiers
Hearing Aids



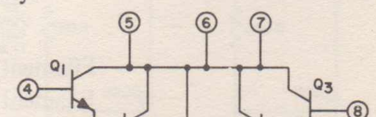
Schematic Diagram for CA3035

CA3036

The CA3036 is a dual Darlington Array having two independant low noise wide band amplifier channels. This circuit is particularly useful for preamplifier and low level amplifier applications in both single channel and stereo systems. It also has a wide application in low noise industrial instrumentation amplifiers.

Features of the CA3036 are:

- Matched low noise transistors
- Darlington connected-emitter follower outputs.
- 200 MHz GBW product.
- High H_{FE} – 4540 typical for Darlington pair.
- Uncommitted output for maximum circuit flexibility.



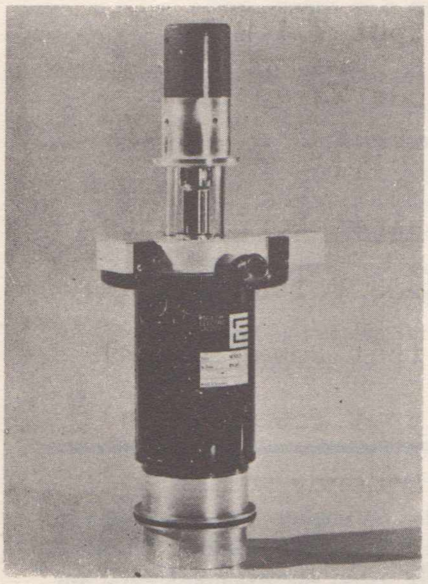
Schematic Diagram for CA3036

A.W.V. SILICON PLANAR EPITAXIAL TRANSISTORS

Elsewhere in this issue data is published on a range of seven new n-p-n silicon epitaxial transistors which are now being manufactured by A.W.V.

The AS300, AS301 and AS302 are designed for general high frequency applications e.g. amplifiers, mixers or oscillators whilst the AS310, AS311, AS312 and AS313 are designed for audio frequency applications.

A feature of these transistors is their epoxy package with a metal can.



EEV M5028 MAGNETRON

The English Electric Valve Co. Ltd. have recently announced a new Magnetron, type M5028 which has been specifically designed for linear accelerators.

The M5028 is a 5 Megawatt, tunable, S-Band, Long-anode Magnetron and has a number of special features:—

i) Integral water jacket to ensure low frequency drift.

ii) Low torque tuning mechanism, incorporating a balanced bellows system, which allows automatic frequency control with low servo effort.

iii) Sturdy construction to minimise both frequency and power variations when the magnetron is orbited about a horizontal axis.

A.W.V. 6166A

In line with its progressive policy of continual development A.W.V. are now manufacturing the first metal ceramic beam power tetrode to be made in Australia. The tube 6166A is the ceramic version of the 6166 which has been manufactured at A.W.V.'s Rydalmere plant for several years.

The 6166A is a forced air cooled tube designed for VHF service in television applications. It has a maximum plate dissipation of 12 Kw and can be used with full ratings up to 220 MHz.

Brief electrical specifications of the 6166A up to 220 MHz are as follows:

Filament Voltage	:	5 ± 5% V
Filament Current	:	168 A
Maximum Plant Voltage	:	7500 V
Maximum Plate Current	:	3 A
Maximum Plate Input	:	20,000 W
Maximum Plate Dissipation	:	12,000 W
Maximum Grid 1 Current	:	0.6 A

VHF Mixer Design

Using the RCA-3N128 Mos Transistor

by

F. M. CARLSON

The 3N128 is a vhf MOS field-effect transistor suitable for use throughout the vhf band (30 to 300 MHz) as an amplifier, mixer, or oscillator. This Note discusses some of the design criteria pertinent to the construction of MOS mixers, and presents an example of a complete vhf MOS converter.

Mixer Design Considerations

The conversion gain obtained from a mixer is the ratio of intermediate-frequency (if) power output divided by the radio-frequency (rf) power input. This conversion gain CG is usually expressed in dB, as follows:

$$CG = 10 \log \frac{\text{if P out}}{\text{rf P in}}$$

The value of CG approximates the gain of the active device operated as an amplifier (unneutralized) at the intermediate frequency, minus the rf losses at the input of the device. Practical mixers normally have a conversion gain of 3 to 5 dB less than their if-amplifier gain.

The 3N128 transistor has good gain and noise figure throughout the vhf band. Because it also has a nonlinear region of operation, it may be used as a vhf mixer to provide good conversion gain. The transfer function of the 3N128, shown in Fig. 1, indicates that the maximum nonlinearities occur at a drain current of about 1.5 milliampere. At drain currents above approximately 5 milliamperes, the transfer function starts to become linear. No mixing action can occur if the transfer function is perfectly linear. Because the amplifier gain of the 3N128 is higher at 5 milliamperes than at 1.5 milliamperes, the best bias point for an MOS mixer is a compromise between the region where best mixing occurs and the region where optimum if power gain occurs. For the 3N128, this point is empirically determined to be between 3.5 and 4.5 milliamperes.

The local-oscillator signal may be introduced into a 3N128 mixer at the insulated gate, the source, or the junction gate (substrate). Application of the oscillator signal to the junction gate can be very effective, but is not recommended because the junction gate of the 3N128 is tied to the case; placing the transistor case at local-oscillator signal potential can pose possible radiation problems.

Injection of the oscillator signal at the source is not desirable because the source should always be at rf/if ground for optimum gain, but could not be at ground at the oscillator frequency. Designing a network to meet these criteria is difficult and adds to the cost of the mixer.

Injection of the oscillator signal at the insulated gate is the least troublesome of the three methods. The local oscillator may be coupled to the insulated gate by means of an inductive loop or a small coupling capacitor.

The input circuit is normally designed for a conjugate match with the input impedance of the MOS transistor at the radio frequency. The output circuit is normally designed for a conjugate match with the output impedance of the 3N128 at the intermediate frequency, unless electrical instabilities (oscillations) occur, in which case the output circuit must be mismatched. Oscillations are not

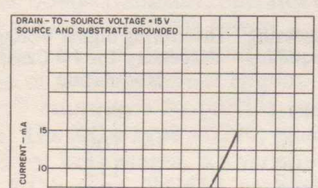


Fig. 1 - Drain current as a function of gate-to=source voltage for the 3N128 vhf MOS transistor.

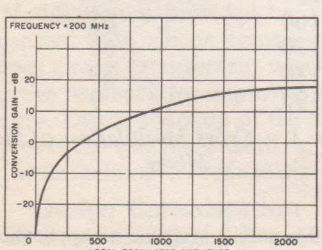


Fig. 3: Conversion gain of the mixer stage in Fig. 2 as a function of local-oscillator amplitude.

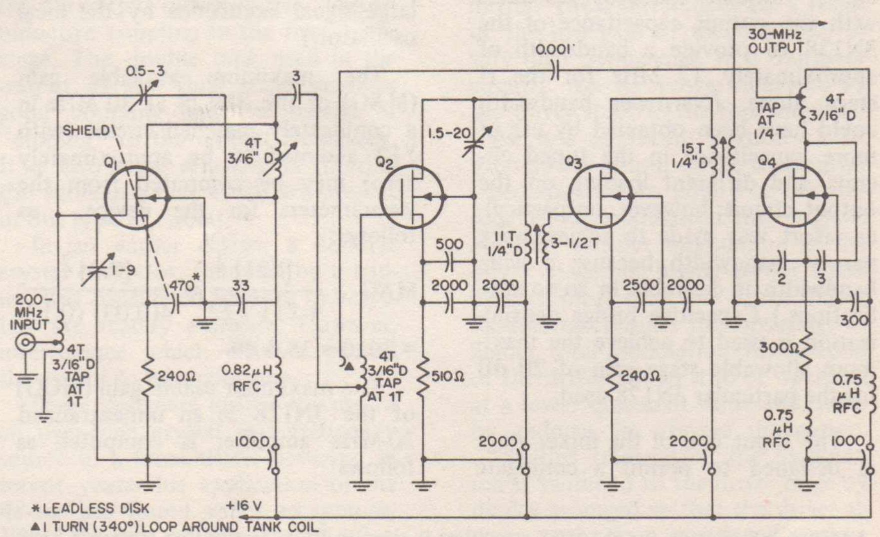


Fig. 2 - VHF receiver "front end" using the 3N128 in all stages.

normally a problem in a 3N128 mixer, provided the if and rf signals are relatively far apart in frequency. Under these conditions, the output circuit presents a low impedance to the rf signal, and the input circuit presents a low impedance to the if signal; consequently, oscillations at either frequency are unlikely to occur.

Neutralization is not generally used in mixers because of the different frequencies at the output and the input.

Design Example

A vhf receiver "front end" has been designed and built to demonstrate the preceding design considerations and to illustrate the use of the 3N128 vhf MOS transistor in all four stages: rf, mixer, if, and local oscillator. The complete converter, shown in Fig. 2, uses an rf input frequency of 200 MHz and an if output frequency of 30 MHz.

The input stage is a straight-through 200 MHz amplifier* employing a source resistor for gate bias. This configuration permits the gate to be at dc ground, and greatly reduces the possibility of damage to the MOS gate from input transients. The 240-ohm resistor allows a current of approximately 5 milliamperes to flow through the device so that maximum vhf power gain is obtained. A variable inductor resonates with the output capacitance of the 3N128 to provide a bandwidth of approximately 12 MHz for the rf stage alone. (Narrower bandwidth could have been obtained by use of more capacitance in the tuned circuits and different loading on the output circuit; however, no particular effort was made to achieve very narrow bandwidth because a wide bandwidth is desirable in some applications.) Capacitive bridge neutralization is used to achieve the maximum allowable stage gain of 20 dB for the particular 3N128 used.

The input coil of the mixer stage is designed to permit a conjugate match with the transistor input admittance. The input admittance y_{11} of the 3N128 at 200 MHz is approximately 0.45 + j 7.2 millimhos. Therefore, an admittance of y_{11}^* (=0.45 – j 7.2 millimhos) should be presented to the mixer input. A conjugate match is not used at the output because it is desirable to load the input of the following (if) stage for stability. Therefore, a step-down transformer is used. (A conjugate match would require use of a step-up transformer because $g_{11} > g_{22}$ at 30 MHz.)

The local oscillator is coupled into the insulated gate of the mixer by means of an ungrounded 340-degree loop placed around the input tank coil at its high-impedance end. A small coupling capacitor could also be used for this purpose. Local-oscillator amplitude is approximately 1.4 volts rms into the coupling loop. Power gain of the mixer stage is 16 dB. Fig. 3 shows the conversion gain of the mixer as a function of local-oscillator amplitude. A lower oscillator level than 1.1 volts would probably be desirable if spurious output frequencies were found to be troublesome.

A 510-ohm source resistor is used in the mixer stage to provide a drain current of 4 milliamperes. (This value is larger than would normally be expected for a drain current of this level because the gate is subjected to large signal excursions by the local oscillator.)

The maximum available gain (MAG) of the 3N128 at 30 MHz in a conjugately matched circuit (with y_{12} assumed to be approximately zero) may be computed from the y-parameters for the device, as follows:

$$\text{MAG} = \frac{[y_{21}]^2}{4\, g_{11}\, g_{22}} = \frac{[7.2]^2}{4(0.03)\,(0.12)}$$

$$= 3610 = 35.6 \text{ dB}$$

The maximum usable gain (MUG) of the 3N128 in an unneutralized 30-MHz amplifier is computed as follows:

$$\text{MUG} = \frac{0.4 g_m}{\omega\, C_{rss}} = \frac{(0.4)(7.2)(10^{-3})}{(2\pi)(30)10^6(0.13)10^{-12}}$$

$$= 118 = 20.7 \text{ dB}$$

If the 30-MHz stage is operated at a gain significantly above the MUG value, the possibility of circuit oscillation exists. Therefore, the input of the if stage is mismatched to reduce the stage gain from the MAG level to about 20 dB. (In some cases it may be easier to mismatch at the output than at the input.) The output of the if stage is transformer-coupled to a 50-ohm load.

The local-oscillator stage is a Colpitts circuit in which frequency is adjusted by means of a slug-tuned inductor.

The complete converter has a power gain of 56 dB, a noise figure of 3 dB, and a bandwidth of 1.5 MHz. Spurious responses (referred to a level of 0 dB at the 200-MHz input frequency) are -51 dB at 100 MHz, -36 dB at 215 MHz, and -35 dB at 260 MHz. These spurious responses could be greatly improved by use of a narrower bandwidth in the rf stage; the present bandwidth is 12 MHz at the 3-dB points.

Table 1 shows measurements of cross-modulation of the converter as a function of frequency.

Interfering Frequency (MHz)	Interfering Signal Voltage Necessary for 1% Cross-Modulation (mV)
50	160
90	115
100	87
110	98
120	97
130	87
140	78
150	77
160	65
170	62
180	39
190	9

Table 1 – Cross Modulation of MOS Mixer.

* Design information for vhf MOS amplifiers is given in RCA Application Note AN-3193: A copy of this note is available on application.

WITH ACKNOWLEDGEMENTS TO R.C.A.

A Solid-State A-M Transmitter for Two-Meter Operation

By D.W. Nelson, WB2EGZ

RCA CAMDEN N.J.

With the advent of RCA "overlay" technology, the useful power and frequency of silicon transistors have been extended to ranges earlier thought unattainable. Today's devices can provided a minimum of 15 watts of output power at 400 MHz or 5 watts at 1 GHz. Simultaneously, continual refinements in development and manufacturing have brought these transistors within reach of the amateur's purse. In the following article, WB2EGZ describes the use of the moderately priced RCA-40290 "overlay" transistor in conjunction with several other transistor types in a 2-meter A-M transmitter having more than 1.5 watts of output power. Although WB2EGZ's transmitter design is most suitable for amateur application, and its construction poses an interesting challenge to the able craftsman, the project is not recommended for the inexperienced builder.

Circuit Description

Figure 1 shows the circuit schematic for this solid-state transmitter. The first three RF stages use silicon diffused transistors. An RCA type 40080 is used for the 72-MHz oscillator stage. The transistor for the second stage, a doubler, is an RCA type 40404 operating, in this case, class A. Although, traditionally, a doubler operates class C, the class C circuit appeared to load the oscillator too much and was unstable during variations in supply voltage. The class A circuit not only eliminates the need for a buffer stage, but provides higher output. The third stage, another 40404 transistor, operates class C as the first 144-MHz driver.

The second driver and the final stage of this transmitter employ "overlay" transistors to provide greater gain. The RCA type 40290, which is used for both stages, is a variation of the RCA type 2N3553 particularly suited for AM systems using low voltage (12-15 volt) power supplies. The 2N3553 would be a good alternate in this circuit, but may have lower power output.

Modulation is provided by six inexpensive RCA silicon transistors in a novel circuit arrangement which has negative feedback for improved audio quality. Direct coupling of the stages improves the efficiency of the amplifier while helping to reduce its cost.

Design Consideration

In a transistorized RF amplifier, one of the greatest problems is that of obtaining an optimum power transfer from one stage to the next. Because of the relatively low gain of the transistors, closer coupling is required for greater power transfer. Unfortunately, the higher efficiency associated with close coupling also results in a wider bandwidth and poorer filtering of the harmonics. Another disadvantage of a close coupling technique is the lack of a "tank" tuned circuit for dipping a Grid Dip Oscillator. Thus a compromise coupling is desirable.

A 72-MHz crystal which operates at the fifth harmonic of its basic frequency helps to reduce the number of troublesome harmonics which are present. A compromise in power was made by the use of a high-Q tank for the output of the doubler, and an inductive coupling to the first driver stage. The double tank used in the output of the final stage provides good coupling and the means of "dipping" to the correct frequency. It is also more effective in filtering the sub-harmonics which are present in this type of circuit.

In an earlier design, a 48-MHz crystal oscillator, followed by a tripler, was considered because the crystals are readily available. However, interference which resulted proved the circuit to be a poor one.

Some designers dislike the use of an over-tone crystal as a frequency source in a transmitter; however, in recent years this application of the device has found wider acceptance. These crystals, which are ground flat, may be excited in the odd harmonics (3rd, 5th, 7th) of the basic frequency by the use of a high-Q circuit tuned to the desired harmonic. The principal disadvantage of an overtone crystal is the low power level at which it must be driven. If the crystal is overdriven to produce the desired output, spurious outputs at slightly higher than the desired frequency will occur. Also, a frequency shift, sometimes permanent, may result. Obviously, either condition is undesirable. However, sufficient amplification, provided by an additional amplifier, avoids these difficulties and establishes the overtone crystal oscillator as a desirable frequency/power source.

Modulation of a transistorized transmitter differs in several ways from that of the normal tube transmitter. These differences are a result of the low gain of the transistor at very high frequencies. In a transistor circuit, the RF power delivered by the driver stage to the final amplifier stage is a significant portion of the total power output of the transmitter —perhaps 10% to 20%. For this reason, 100% modulation may not be possible without modulation of the driver stage. In higher-power transmitters, more than one driver will be modulated. Besides the advantage of higher total modulation, modulation of the driver allows it to be operated at a lower quiescent value and thereby reduces its average dissipation.

In this transmitter, the modulation is switched to the driver stage by diodes arranged so that the driver always peaks in a positive direction. This technique has been found very useful in assuring upward modulation.

Transistor Cooling Considerations

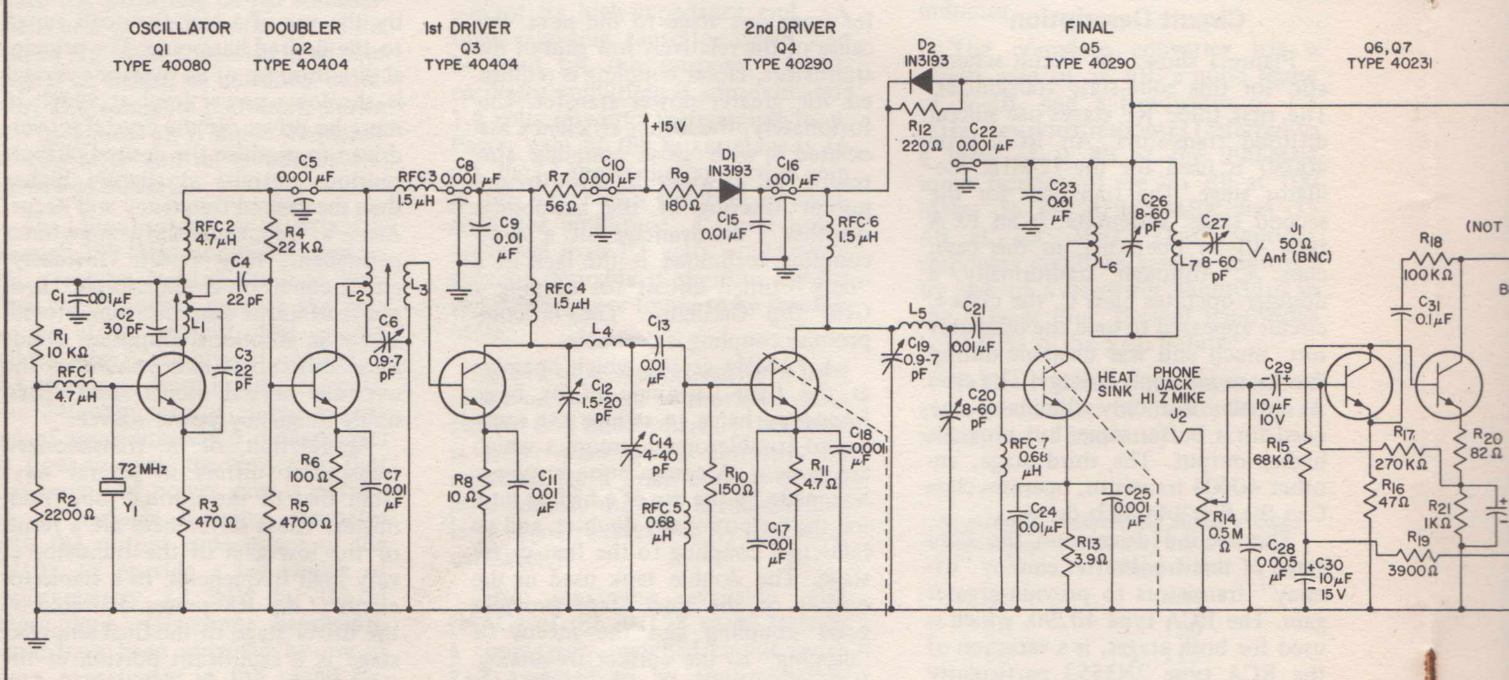
The power rating of a transistor or diode is based on the maximum safe operating temperature of the semiconductor junction. In all power applications, transistors must be properly cooled by conduction of heat away from the junctions. In this transmitter, cooling of the RF final stage is provided by a special heat sink; cooling of the modulator output transistors is effectively provided by the brass chassis. Thermal conductivity from the transistor to the heat sink is improved by the application of a thermal compound, such as silicone grease. Figure 2 illustrates how the modulator output transistors are insulated from the chassis by means of a mica spacer and two nylon feed-through washers supplied with the transistor. Although silicon semiconductors can withstand higher temperatures than germanium types, *it cannot be stressed too strongly that heat sinks and a thermal compound must be used to achieve long life in power transistors.*

Furthermore, transistors are easily destroyed by momentary overloads; *therefore, connections should be double-checked before the transistors are inserted in the sockets.* Also, remember that *transistors should never be inserted or removed when the power is on.*

If, for some reason, a stage is unstable and oscillates, the transistor may become overheated. It is wise, therefore, to keep a receiver turned on while experimenting with the transmitter. If a stage breaks into oscillation, noise will be heard over a wide band on the receiver. At this point, turn off the transmitter, correct the problem, and be sure that the transistors are reasonably cool before applying power again.

Construction

The entire transmitter is assembled on a sheet of brass measuring 5 inches by 9½ inches. This is later mounted on an inverted 5-by-9½-by-



C_1, C_7, C_9, C_{11}, C_{13}, C_{15}, C_{17}, C_{21}, C_{23}, C_{24} — 0.01 μF, 50 volt, ceramic disc
C_2 — 30 pF, mica
C_3, C_4 — 22 pF, mica
C_5, C_8, C_{10}, C_{16}, C_{22} — 0.001 μF Filtercon or equiv.
C_6, C_{19} — 0.9-7 pF mica variable (Arco 400 or equiv.)
C_{12} — 1.5-20 pF mica variable (Arco 402 or equiv.)
C_{14} — 4-40 pF mica variable (Arco 403 or equiv.)
C_{18}, C_{25}, C_{37} — 0.001 μF, 50 volt, ceramic disc
C_{20}, C_{26}, C_{27} — 8-60 pF mica variable (Arco 404 or equiv.)
C_{28} — 0.005 μF, 50 volt, ceramic disc
C_{29} — 10 μF, 10 volt Tantalum or equiv.
C_{30} — 10 μF, 15 volt Tantalum or equiv.
C_{31}, C_{36} — 0.1 μF, 100 volt, paper
C_{32} — 200 μF, 3 volt Tantalum or equiv.
C_{33}, C_{35} — 100 μF, 20 volt Tantalum or equiv.
C_{34} — 50 μF, 6 volt Tantalum or equiv.
L_1 — 4 turns, No. 18 wire, tapped ½ turn and 1 turn from cold end, wound on ¼-inch ceramic form, Green Core
L_2 — 4 turns, No. 22 wire, tapped 1 turn from hot end, wound on ¼-inch ceramic form, White Core
L_3 — 2 turns, No. 22 wire on same form as L_2, close to brass chassis
L_4, L_5 — 3 turns, No. 16 wire, ¼-inch inside diameter
L_6 — 4 turns, No. 16 wire, tapped 1½ turns from cold end, ¼-inch inside diameter
L_7 — 4 turns, No. 16 wire, tapped 1¾ turns from cold end, ¼-inch inside diameter
RFC_1, RFC_2 — 4.7 μH choke
RFC_3, RFC_4, RFC_6 — 1.5 μH choke
RFC_5, RFC_7 — 0.68 μH choke
R_1 — 10 kilohms, ¼ watt
R_2 — 2,200 ohms, ¼ watt
R_3 — 470 ohms, ¼ watt
R_4 — 22 kilohms, ¼ watt
R_5 — 4,700 ohms, ¼ watt
R_6 — 100 ohms, ¼ watt
R_7 — 56 ohms, ½ watt
R_8 — 10 ohms, ¼ watt
R_9 — 180 ohms, ½ watt
R_{10} — 150 ohms, ¼ watt
R_{11} — 4.7 ohms, ½ watt
R_{12} — 220 ohms, ½ watt
R_{13} — 3.9 ohms, ½ watt
R_{14} — 0.5 megohm potentiometer
R_{15} — 68 kilohms, ¼ watt
R_{16} — 47 ohms, ¼ watt
R_{17} — 270 kilohms, ¼ watt
R_{18} — 100 kilohms, ¼ watt
R_{19} — 3,900 o
R_{20} — 82 ohms
R_{21} — 1,000 o
R_{22} — 1,000 o
R_{23} — 10 kiloh
R_{24} — 1,500 o
R_{25} — 2.5 ohm wound
T_1 — Stancor T
T_2 — Stancor T
Y_1 — 72 MHz Crystal Type

Miscellaneous ets (Augat 5 transistor 3301 or e

Figure 1: Schematic diagram and parts list of the WB2EG

2-inch chassis as shown in Figure 2. Figures 3 and 4 show the top and bottom views, respectively, of the assembled transmitter. Although the brass sheet has been silver-plated for enhanced electrical conductivity, the transmitter should operate satisfactorily without the plating. No attempt has been made to compress the size of the transmitter. Leave the task of miniaturization to the professionals.

Transistor sockets have been used, except with the audio power units, to eliminate the danger of overheating the leads when soldering. The use of sockets also facilitates replacement of the transistors when necessary. Although small Augat sockets were used for the author's circuit, these may be replaced by the TO-5 sockets without sacrificing efficiency. TO-5 sockets were used on the breadboard and are somewhat easier to solder.

The greatest single cause of poor transmitter efficiency is ineffective bypassing. The best bypass capacitors are feed-through types having a ferrite filtering element. An improvement over the single bypass element may be obtained by the use of a second capacitor having a different value. Ceramic disc capacitors are effective in bypassing and have been used wherever practical to reduce cost.

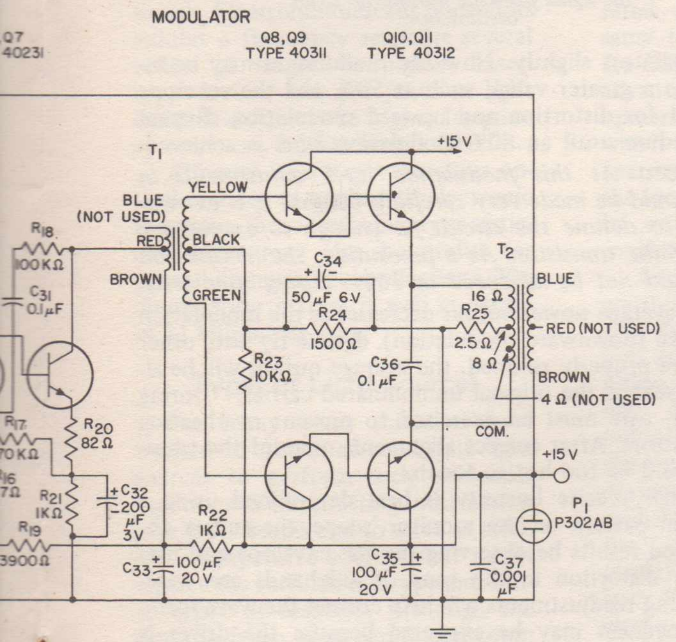
Alignment

The suggested steps for alignment are as follows:

1. With the RF transistors secured in their sockets and the power off, tune the oscillator, doubler, and final tank circuits to their respective frequencies.

2. Remove the 40290's. apply power, and tune the first three stages for a maximum current flow by measuring the voltage drop across the 56-ohm resistor (R_7). This reading should be about 2.0 volts. At this moment, the signal heard on your receiver should be very strong.

3. Next, connect the output of the transmitter to an appropriate 50-ohm load and insert the 40290 transistors, remembering that power must be off when inserting or removing transistors. Assuming that the circuit



R_{19} — 3,900 ohms, ½ watt
R_{20} — 82 ohms, ¼ watt
R_{21} — 1,000 ohms, ¼ watt
R_{22} — 1,000 ohms, ½ watt
R_{23} — 10 kilohms, ½ watt
R_{24} — 1,500 ohms, ½ watt
R_{25} — 2.5 ohms, 3 watts, wire-wound
T_1 — Stancor TA31 or equiv.
T_2 — Stancor TA56 or equiv.
Y_1 — 72 MHz crystal, International Crystal Type FA5, or equiv.

Miscellaneous — 4 transistor sockets (Augat 8060-1G5 or equiv.); 5 transistor sockets (Elco 05-3301 or equiv.); 1 heat sink (Thermalloy 2207-PR10 or equiv.); 1 aluminum chassis (5 by 9½ by 2 inches); 1 brass plate (5 by 9½ by 1/16 inches); 1 crystal socket (Elco 04-210-04 or equiv.); 1 BNC connector; 1 phone jack; 1 connector (Cinch-Jones P302AB or equiv.)

Note: All resistor power ratings are recommended minimum values.

f the WB2EGZ solid-state A-M transmitter for two-meter operation.

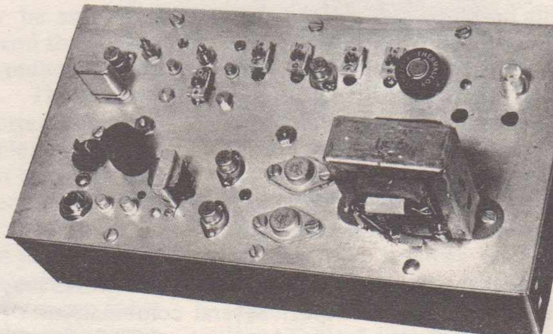


Figure 2: Assembled transmitter, showing RF section at upper left and modulator section at lower right.

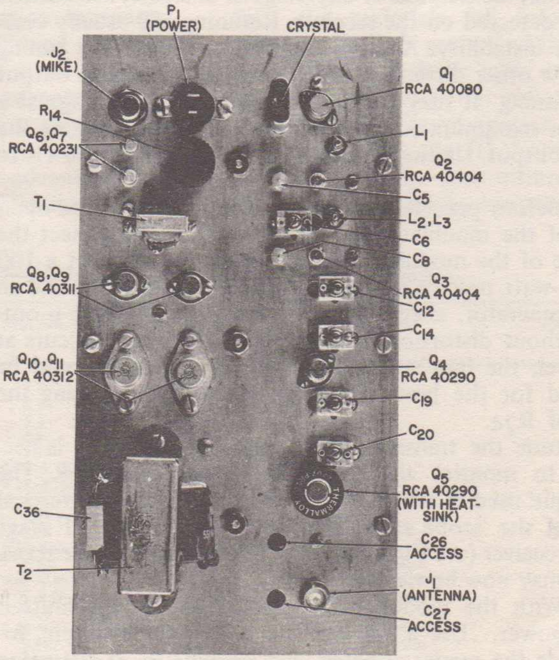


Figure 3: Top view of transmitter.

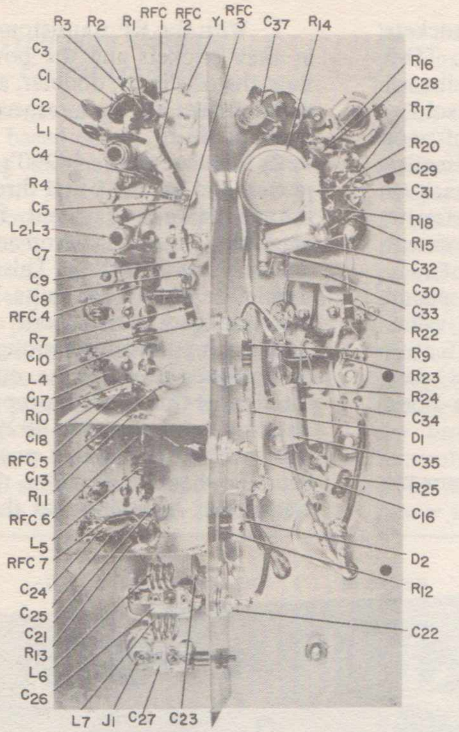


Figure 4: Bottom view of transmitter, showing RF section above and modulator section below.

is stable when the power is turned on, the pi networks may be tuned for miximum output. Several combinations of capacitor settings may give good output, so look for the best combinations of tuning. Retuning the oscillator and buffer may be helpful. If any stage is unstable, oscillations will be detected on the receiver. Retuning will usually correct this instability. Another indication of improper tuning (or some other difficulty) is a very abrupt change in output when tuning. It may be necessary to tune one stage at a time for maximum collector current before optimizing the power output. Unmodulated power output should be 1½ to 2 watts.

4. Before proceeding with the alignment, proper operation of the modulator should be verified. Disconnect the RF side of the modulation transformer and connect a 100 ohm, 1-watt resistor across it. Using an oscilloscope and an audio generator, check the wave-form for a 28-V p-p output without distortion. If clipping or distortion occurs at this level, the bias voltage for the power stage should be adjusted for the best sinusoidal pattern by changing the value of R_{24}.

To tune the transmitter for best modulation, it is necessary to monitor the signal with an oscilloscope. The technique used in this instance was to connect the vertical input of the 'scope across the output of the final IF stage in the receiver (see Figure 5.) The RF envelope of the transmitter may now be viewed.

5. With the modulator connected to the transmitter, apply power. Use a test tone, a 1 KHz for example, to modulate the carrier at some level below 30%. If distortion is observed in the envelope, adjust the final tank circuit and

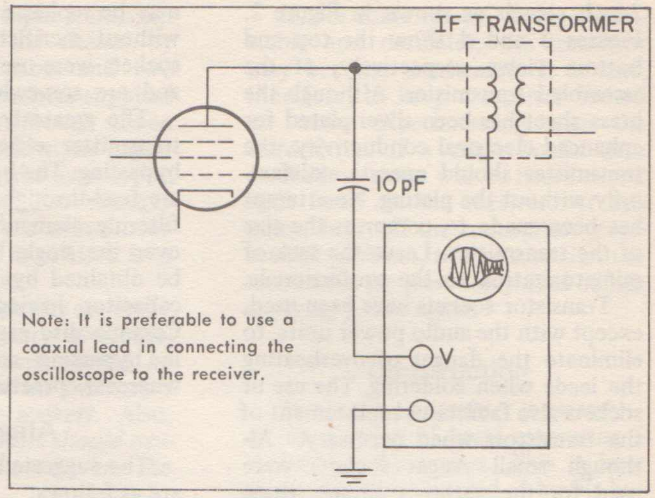


Figure 5: Schematic diagram showing method used by WB2EGZ to connect the signal from the transmitter to an oscilloscope.

input capacitors slightly. Now the modulation may be increased to a greater value, such as 50%, and the envelope rechecked for distortion and upward modulation. Repeat the procedure until an 80% modulation level is achieved.

Caution: *At this modulation level, adjustments in tuning should be made very carefully because it is extremely easy to detune the circuit far enough to exceed the ratings of the transistor. As a precaution, the modulation level should not be advanced to 100% during tone tests.*

If the average power output decreases as the modulation is increased (downward modulation), detune L_2 until other circuits are properly retuned, the average output will be almost as great as the original unmodulated carrier. During alignment, care must be exercised to prevent overheating the transistors. After correct alignment, none of the transistors should be too hot to touch.

Although precise linearity is best determined using a trapezoidal pattern on the monitor 'scope, the author obtained good results be observing the RF envelope and listening for distortion in the tone. If sidebands are wide, slight tuning readjustments will help correct the wave form. Some broadness may be expected because the driver is modulated and because the output capacitance changes when a transistor is modulated. Before going on the air, set the modulation level to peak at 90% with the microphone to be use.

"On the Air"

The power output averages 1.5 watts when using a 15-volt supply. Power consumption at this level is about 8 watts. Although operation of the transmitter from a 12-volt source is possible, the power output will be reduced. Should you desire to operate the transmitter using a power supply having wide variations of output voltage, it is advisable that you tune the transmitter using the lowest supply voltage. Direct operation from an automobile supply would not be practical without providing some means of regulation within the transmitter.

Note: An actual size chassis template is available on request.

P-N JUNCTIONS AS OPTICAL SOURCES

M.F. LAMORTE

R.C.A. SOMERVILLE, N.J., U.S.A.

At present, the most widely used types of solid-state optical sources are the tungsten filament and the electroluminescent phosphors employed in cathoderay tubes and television screens. The former is a black-body radiator and uses only a portion of the continuous spectrum in any one applications. In the latter an electron beam is used to excite the phosphor, and a series of intense lines is produced rather than a continuum. The tungsten lamp has found little application in electronic functions because of the poor frequency response of its radiation to high-frequency electrical signals. Electroluminescent phosphors exhibit a frequency response several orders of magnitude higher. Laser-type devices and p-n junction light emitters exhibit even higher frequency response, approaching 1 GHz.

Although this paper discusses optical sources as applied to electronic functions, these electronic-type optical sources may also be useful for general illumination applications. The illumination market is well in excess of $1 billion per year.

Types of Light Sources

For many years scientists have worked to obtain all-solid-state light sources to perform such electronic functions as transmitting information over light beams and to serve as efficient, inexpensive sources that might replace the cathode-ray type and other types of displays. Although the electron beam as an exciting medium offers some attractive features, low-voltage operation and flat or mural-type display are desirable at price and performance levels which would be competitive with present-day devices.

One type of all-solid-state source which has been under intensive study is the high-field AC and DC electroluminescent device. At the present stage of development, these devices have the disadvantages of low efficiency, poor operating-life characteristics, and high-voltage operation (which makes them incompatible with silicon transistors). If improvements can be made in these areas, however, this type of device would be attractive in some applications for mural displays.

Perhaps the most attractive all-solid-state light source is the p-n junction luminescent diode. Diodes fabricated from many of the varied crystals that exhibit electroluminescence may also be placed in the stimulated-emission mode. The use of this device provides the advantages of high external efficiency in both the incoherent and coherent modes, long operating life, and compatibility with driving circuit employing silicon devices. In addition, the high external efficiency reduces the driving-power requirements, and may make possible the use of low-cost silicon integrated circuits. In some cases, the same technology may be employed for different crystals, and the further advantage of greatly reduced development cost results. The diodes which have been reported cover the near-infrared and visible portions of the spectrum. The light intensity may be modulated at frequencies approaching 1 GHz.

Applications for p-n junction light emitters range from electroluminescent displays to such electronic functions as card reading, character recognition, sensing, electro-optical switching, optical ranging, illumination, metrology, communication, intrusion alarms, control circuits, and warning devices. The non-laser diode appears to have a substantial advantage over the laser in present-day devices for display applications; however, the laser diode has distinct advantages for electronic-function applications.

In the following section, the nature of p-n junction luminescence is discussed, and the various crystal materials used in light-emitting diodes are classified according to the portion of the spectrum in which each emits energy.

Light-emitting Diodes

In general, light is emitted from a p-n junction when the diode is forward-biased. Minority carriers are injected into the opposite-conductivity -type material and a nonequilibrium condition is created which causes the carriers to recombine radiatively. External efficiency is proportional to the injection efficiency of the junction. Because the injection efficiency may approach unity when the junction is properly designed, the external efficiency may also approach unity.

There are many known semiconductor materials, and probably many unknown at this time, which emit light when properly fabricated into diodes. Fig. 1 shows various crystal materials and the portion of the spectrum in which they emit radiation. Table 1 lists these materials and indicates those which also provide laser action. With the exception of SiC. the materials fall into three groups: the III-V group and the II-VI group of the periodic table and lead salts. The lead salts emit radiation toward the far-infrared portion of the spectrum, while the III-V and II-VI groups cover the near-infrared and visible portions of the spectrum. The

TABLE I—Listing of p-n Junction Light Emitting Diodes

Crystal	*Macrometers*	*Laser Action*	*Remarks*	*References*
PbSe	8.5	Yes	direct	2
PbTe	6.5	Yes	direct	2, 3
InSb	5.2	Yes	direct	4
PbS	4.3	Yes	direct	2
InAs	3.15	Yes	direct	5
$(In_x Ga_{1-x}) As$	0.85-3.15	Yes	direct	6
$In (P_x As_{1-x})$	0.91-3.15	Yes	direct	7
GaSb	1.6	No		8
InP	0.91	Yes	direct	9
GaAs	0.90	Yes	direct	10, 11, 12
$Ga (As_{1-x} P_x)$	0.55-0.90	Yes	direct	13
CdTe	0.855	No	homojunction	14
$(Zn_x Cd_{i-x}) Te$	0.59-0.83	No	homojunction	15
$CdTe\text{-}Z_oTe$	0.56-0.66	No	heterojunction	16
BP	0.64	No	indirect	17
$Cu_2Se\text{-}ZnSe$	0.40-0.63	No	heterojunction	18
$Zn (Se_x Te_{1-x})$	0.627	No	homojunction	19
ZnTe	0.62	No	barrier	20
Ga P	0.565	No	indirect-band gap	21
	0.68	No	indirect-oxygen line	
Si C	0.456	?	-SiC homojunction	22

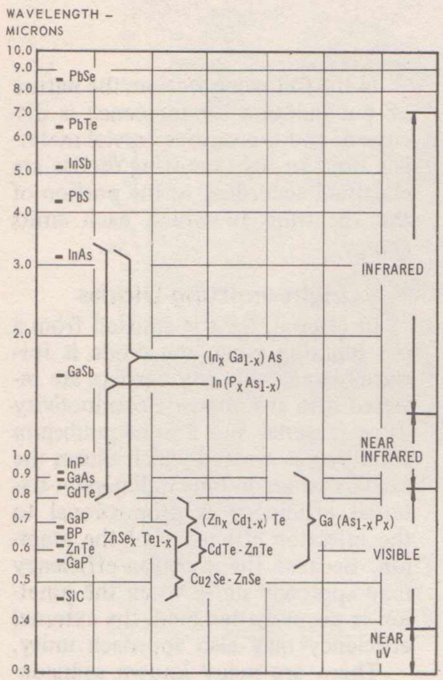


Fig. 1 – Various crystal materials and the portion of the spectrum in which they emit radiation.

latter two groups exhibit considerable overlap. Continuous coverage of the spectrum from the blue to the near-infrared region can be obtained by use of mixed crystals.

Generally, any one diode emits one strong line that has width in the order of 300 angstroms. The portion of the spectrum in which the diode emits depends on its forbidden-band energy and/or the position that the dopant takes in the forbidden band. Each semi-conductor, including the mixed crystals, possesses a unique bandgap energy and emits a charactistic wavelength. With the crystals shown in Table 1, the entire spectrum from 0.45 to 3.1 micrometers can be covered. Further investigation will probably extend this coverage out to 8.5 micrometers. If the II-VI compounds do not adequately cover the spectrum in the blue and near-ultraviolet regions, another group of materials will have to be investigated for this region.

GaAs light-emitting diodes have received more attention than other types for several reasons.

1) Existing military and industrial applications require this type of infrared-emitting device.

2) The emission of GaAs devices matches the peak response of both silicon detectors and photodevices that have an S-1 response.

3) More advanced technology for GaAs crystal growth and device fabrication existed at the time that the most intensive studies on p-n junction light sources were conducted.

4) An ample supply of high-quality GaAs crystal is available from commercial establishments.

5) A certain amount of confidence in fabrication techniques exists as a result of previous experience on other GaAs devices.

In addition, many of the problems associated with increasing the external efficiency of p-n junction light sources are similar to those encountered on other GaAs devices. Therefore, GaAs affords a convenient vehicle for investigation of these problems and improvement of light-emitting devices.

Performance of P-N Junction Diodes

The performance of a light-emitting device is determined in large measure by the quality of the p-n junction. It is difficult to form high-injection-efficiency junctions in some crystals. In the lead salts and the III-V compounds, both n- and p-type crystals may be obtained easily; therefore, homojunctions of high quality may be obtained. In the II-VI compounds, however, n- and p-type crystals of the same crystalline material are not always possible; as a result, heterojunctions must be employed. Because heterojunctions are usually inferior to homojunctions, particularly with respect to injection efficiency, the II-VI compounds are used as a last resort.

Crystalline defects are introduced during the construction of heterojunctions as a result of the mismatch (however small) in the crystalline structure of two different types of crystal. These defects produce energy levels in the band gap at the junction region of the diode. In most cases, depending on the defect density and its activation energy, these defect energy levels adversely affect the internal radiative recombination efficiency. The most probable result is that the defect levels cause a nonradiative recombination process which tends to reduce the radiative recombination efficiency. Moreover, heterojunctions usually require more complex technology, are more difficult to control, and are more difficult to obtain reproducibly. If a choice exists, therefore, homojunctions are more desirable.

A disadvantage of using wide-band-gap materials for luminescent diodes is that low contact resistance is more difficult to obtain. These materials (particularly SiC and GaP) cannot be operated at high-current-density values because of their high contact resistance, coupled with the higher thermal resistance of diatomic and ternary compounds. These resistances serve to reduce the brightness of the emitted radiation. In addition, the impurity atoms assume energy levels within the band gap somewhat removed from the band edges, and result in large activation energies. This condition may lead to carrier freeze-out at low temperatures and render the diodes inoperative at such temperatures. However, improvements in technology and crystalling quality may minimize and perhaps even eliminate these problems.

GaAs diodes exhibit higher efficiencies than any other light-emitting diodes. This greater efficiency can be attributed to the higher crystal quality and more advanced fabrication techniques. Improvements must be made in other materials to match the performance of GaAs diodes. In principle, there is nothing to prevent any of the light-emitting diodes from attaining an internal quantum efficiency of unity. The external efficiency is usually low because of reabsorption and a small critical angle in most semiconductor materials.

Editor's Note:
There are three developmental GaAs diodes currently available TA2628, TA2930, TA7008. Details on these devices are available on request.

THEORY AND APPLICATION OF THYRISTORS (PART 2)

CONSTRUCTION

Construction details for typical RCA thyristors are shown in Figs. 8 through 12. Fig. 8 shows details for the 2-lead TO-5 package. This compact package is designed for applications in which mounting space is limited and can be attached to a wide variety of heat sinks with sizes and shapes to fit the available space. A typical heat-sink arrangement for an insulating mounting of this package is shown in Fig. 9.

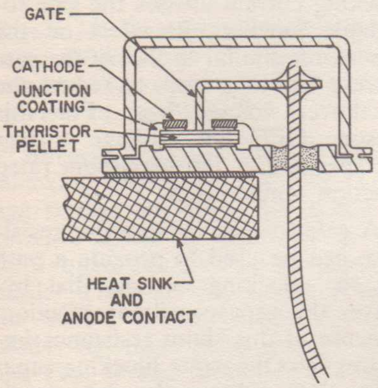


Figure 8. Cross-section of RCA two-lead TO-5 thyristor package.

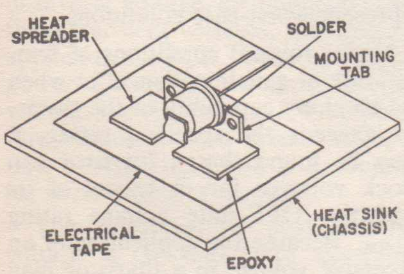


Figure 9. Typical heat-sink isolation technique for a chassis-mounted two-lead TO-5 thyristor.

This package is used at current levels up to 7 amperes.

In higher-current applications the TO-66, TO-3, and press-fit and stud-mounted TO-48 packages are used. Internal construction details of the press-fit package are shown in Fig. 10.

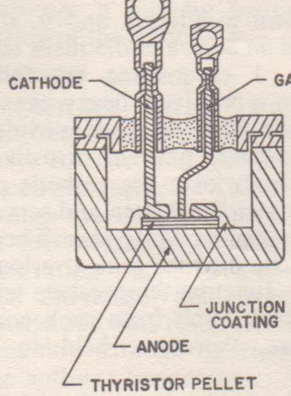


Figure 10. Cross-section of RCA press-fit thyristor package.

Construction details of a typical SCR pellet are shown in Fig. 11. The shorted-emitter construction used in RCA SCR's can be recognized by the metallic cathode electrode in direct contact with the p-type base layer around the periphery of the pellet. The gate, at the centre of the pellet, also makes direct metallic contact to the p-type base so that the portion of this layer under the n-type emitter acts as an ohmic path for current flow between gate and cathode. Because this ohmic path is in parallel with the n-type emitter junction, current preferentially takes the ohmic path until the IR drop in this path reaches the junction threshold voltage of about 0.8 volt. When the gate voltage exceeds this value, the junction current increases rapidly, and injection of electrons by the n-type emitter reaches a level high enough to turn on the device.

In addition to providing a precisely controlled gate current, the shorted-emitter construction also improves the high-temperature and dv/dt (maximum allowable rate of rise of OFF-state voltage) capabilities of the device. The junction depletion layer acts as a parallel-plate capacitor which must be charged when blocking voltage is applied. Because the charging, or displacement, current ($i = C dv/dt$) into this capacitor varies as the rate of rise of forward voltage (dv/dt), a very high dv/dt can result in a high current between anode and cathode. If this current crosses the n-type emitter junction and is of the same order of magnitude as the gate current, it can trigger the device into the conducting state. Such unwanted triggering is minimized by the shorted-emitter construction because the peripheral contact of the p-type base to the cathode electrode provides a large-area parallel path by which the dv/dt current can reach the cathode electrode without crossing the n-type emitter junction.

The centre-gate construction of the SCR pellet provides fast turn-on and high di/dt capabilities. In an SCR, conduction is initiated in the cathode region immediately adjacent to the gate contact and must then

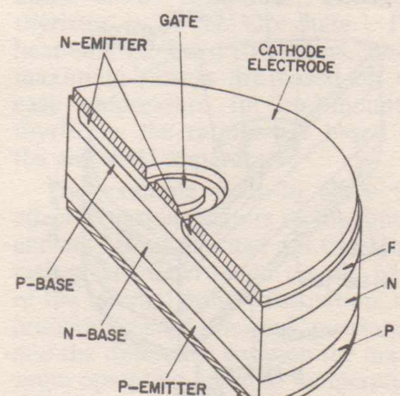


Figure 11. Cross-section of a typical SCR pellet.

propagate to the more remote regions of the cathode. Switching losses are influenced by the rate of propagation of conduction and the distance conduction must propagate from the gate. With a central gate, all regions of the cathode are in close proximity to the initially conducting region so that propagation distance is significantly decreased; as a result, switching losses are minimized.

Construction of a typical RCA triac pellet is shown in Fig. 12. In this device, the main-terminal-No. 1 electrode makes ohmic contact to a p-type emitter as well as to an n-type emitter. Similarly, the main-terminal-No. 2 electrode also makes ohmic contact to both types of emitter, but the p-type emitter of the main-terminal-No. 2 side is located opposite the n-type emitter of the main-terminal-No. 1 side, and the main-terminal-No. 2 n-type emitter is opposite the main-terminal-No. 1 p-type emitter. The net result is two four-layer switches in parallel, but oriented in opposite directions, in one silicon pellet. This type of construction makes it possible for a triac either to block or to conduct current in either direction between main-terminal-No. 1 and main-terminal-No. 2.

The gate electrode also makes contact to both n- and p-type regions. As a result, the device can be triggered by either positive or negative gate signals, for either polarity of voltage between the main-

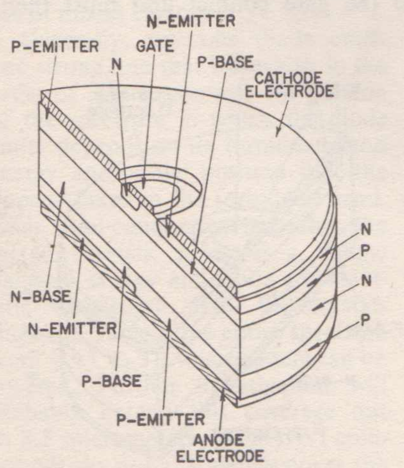


Figure 12. Cross-section of a typical triac pellet.

terminal electrodes. When the triac is triggered by a positive gate signal, conduction is initiated, as in the SCR, by injection of electrons from the main-terminal-No. 1 n-type emitter, and the gate n-type region is passive. The gate n-type region becomes active when the triac is triggered by a negative gate signal, because it then acts as the n-type emitter of a grounded-base n-p-n transistor. Electrons injected from this region enter the n-type base and cause a forward bias on one of the p-type emitters, depending on which is at the positive end of the voltage between the main-terminal electrodes.

As shown in Figs. 8 and 9, the cathode of an SCR and the main terminal-No. 1 of a triac are fully covered by a relatively heavy metallic electrode. This electrode provides a low-resistance path to distribute current evenly over the cathode or main-terminal-No. 1 area and serves as a thermal capacitor to absorb heat generated by high surge or overload currents. Junction-temperature excursions that result from such conditions are, therefore, held to a minimum.

RATINGS AND CHARACTERISTICS

Thyristors must be operated within the maximum ratings specified by the manufacturer to assure best results in terms of performance, life, and reliability. These ratings define limiting values, determined on the basis of extensive tests, that represent the best judgment of the manufacturer of the safe operating capability of the device. The manufacturer also specifies a number of device parameters, called characteristics, which are directly measurable properties that define the inherent qualities and traits of the thyristor. Some of these characteristics are important factors in the determination of the maximum ratings and in the prediction of the performance, life, and reliability that the thyristor can provide in a given application.

VOLTAGE RATINGS

The voltage ratings of thyristors are given for both steady-state and transient operation and for both forward- and reverse-blocking conditions. For SCR's, voltages are considered to be in the forward or positive direction when the anode is positive with respect to the cathode. Negative voltages for SCR's are referred to as reverse-blocking voltages. For triacs, voltages are considered to be positive when main terminal No. 2 is positive with respect to main terminal No. 1. Alternatively, this condition may be referred to as operation in the first quadrant.

OFF-State Voltage—The repetitive peak OFF-state voltage V_{DRM} is the maximum value of OFF-state voltage, either transient or steady-state, that the thyristor should be required to block under the stated conditions of temperature and gate-to-cathode resistance. If this voltage is exceeded, the thyristor may switch to the ON state. The circuit designer should insure that the V_{DRM} rating is not exceeded to assure proper operation of the thyristor.

The effect of increased temperature is accentuated in thyristors because of the regenerative action upon which the operation of these devices is dependent. Thermally generated currents tend to be multiplied. If this blocking current crosses the gate-to-cathode junction, its effect on the thyristor is similar to that of the gate current and thus tends to reduce the breakover voltage V_{BO}. For this reason, OFF-state voltage ratings are specified at the maximum rated junction temperature.

A gate-to-cathode shunting resistance can be used to provide a path for the blocking current that bypasses the gate-to-cathode junction. The use of this shunt resistance improves the OFF-state blocking capability, but reduces the gate sensitivity. OFF-state voltage ratings, therefore, are specified with no external gate-to-cathode impedance to represent worst-case conditions.

Under relaxed conditions of temperature or gate impedance, or when the blocking capability of the thyristor exceeds the specified rating, it may be found that a thyristor can block voltages far in excess of its repetitive OFF-state voltage rating V_{DRM}. Because the application of an excessive voltage to a thyristor may produce irreversible effects, an absolute upper limit should be imposed on the amount of voltage that may be applied to the main terminals of the device. This voltage rating is referred to as the peak OFF-state voltage V_{DM}. It should be noted that the peak OFF-state voltage has a single rating irrespective of the voltage grade of the thyristor. This rating is a function of the construction of the thyristor and of the surface properties of the pellet. The V_{DM} rating should not be exceeded under either continuous or transient conditions.

Fig. 13 shows a simple, inexpensive test circuit that may be used to evaluate the OFF-state voltage capabilities of thyristors. (The circuit may also be used for reverse-blocking and leakage tests of thyristors.) Resistor R_1 and capacitor C_1 are included in the test circuit to limit the rate of rise of applied voltage to the thyristor under test. Resistor R_2 limits the discharge of capacitor C_1 through the thyristor in the event that the thyristor is turned on during the test. Resistor R_3 provides a discharge path for capacitor C_1.

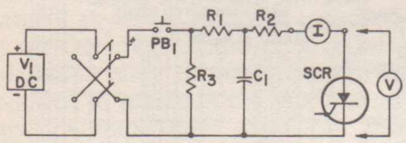


Figure 13. Test circuit used to determine dc forward- and reverse-voltage-blocking capabilities and leakage current of thyristors.

Reverse Voltages (For Reverse-Blocking Thyristors)—Reverse voltage ratings are given for SCR's to provide operating guidance in the third quadrant, or reverse-blocking mode. There are two voltage ratings for SCR's in the reverse-blocking mode: repetitive peak reverse voltage (V_{RRM}) and nonrepetitive peak reverse voltage (V_{RSM}).

The repetitive peak reverse voltage is the maximum allowable value of reverse voltage, including all repetitive transient voltages, that may be applied to the SCR. Because reverse power dissipation is small at this voltage, the rise in junction temperature because of this reverse dissipation is very slight and is accounted for in the rating of the SCR.

The nonrepetitive peak reverse voltage is the maximum allowable value of any nonrepetitive transient reverse voltage which may be applied to the SCR. These nonrepetitive transient voltages are allowed to exceed the steady-state ratings, even though the instantaneous power dissipation can be significant. While the transient voltage is applied, the junction temperature may increase, but removal of the transient voltage in a specified time allows the junction temperature to return to its steady-state operating temperature before a thermal runaway occurs.

The test circuit shown in Fig. 13 may be used for reverse-voltage tests of an SCR.

ON-State Voltage—When a thyristor is in a high-conduction state, the voltage drop across the device is no different in nature from the forward-conduction voltage drop of a rectifier, although the magnitude may be slightly higher. As in rectifiers, the ON-state voltage-drop characteristic is the major source of power losses in the operation of the thyristor, and the temperatures produced becomes a limiting feature in the rating of the device.

CURRENT RATINGS

Thyristor current ratings define maximum values for normal or repetitive currents and for surge or nonrepetitive currents. These maximum ratings are determined on the basis of the maximum junction-temperature rating, the junction-to-case thermal resistance, the internal power dissipation that results from the current flow through the thyristor, and the ambient temperature. The effect of these factors in the determination of current ratings is illustrated by the following example:

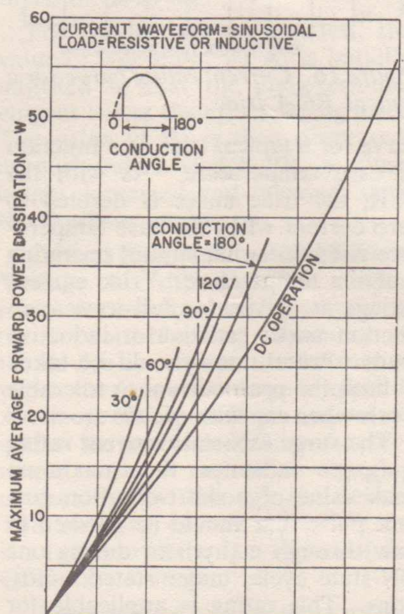


Figure 14. Power-dissipation rating chart for the RCA-2N3873 SCR.

Fig. 14 shows curves of the maximum average forward power dissipation for the RCA-2N3873 SCR as a function of average forward current for dc operation and for various conduction angles. For the 2N3873, the junction-to-case thermal resistance $\theta_{J\text{-}C}$ is 0.92° C. per watt and the maximum operating junction temperature T_J is 100° C. If the maximum case temperature $T_{C(max)}$ is assumed to be 65° C., the maximum average forward power dissipation can be determined as follows:

$$P_{AVG(max)} = \frac{T_{J(max)} - T_{C(max)}}{\Theta_{J\text{-}C}} \quad (6)$$

$$= \frac{(100 - 65)\ ^\circ\mathrm{C}}{0.92\ ^\circ\mathrm{C/watt}}$$

$$= 38 \text{ watts}$$

The maximum average forward current rating for the specified conditions can then be determined from the rating curves shown in Fig. 14. For example if a conduction angle of 180 degrees is assumed, the average forward current rating for a maximum dissipation of 38 watts is found to be 22 amperes.

These calculations assume that the temperature is uniform throughout the pellet and the case. The junction temperature, however, increases and decreases under conditions of transient loading or periodic currents, depending upon the instantaneous power dissipated within the thyristor. The current rating must take these variations into account.

ON-State Current—The ON-state current ratings for a thyristor indicate the maximum values of average, rms, and peak (surge) current that should be allowed to flow through the main terminals of the device, under stated conditions, when the thyristor is in the ON state. For heat-sink-mounted thyristors, these maximum ratings are based on the case temperature; for lead-mounted thyristors, the ratings are based on the ambient temperature.

The example used to show the effect of various factors on maximum current ratings pointed out that these ratings are determined on the basis of the internal power dissipation, the junction-to-case thermal resistance, and the difference between the maximum operating junction temperature and the maximum case temperature. Because the maximum operating junction temperature is fixed, the maximum ON-state current ratings may be given by curves that relate current to case temperature. The maximum allowable current approaches zero as the case temperature approaches the maximum operating junction temperature because this current is directly proportional to the ratio of the difference between case and junction temperatures to the junction-to-case thermal resistance.

The **maximum average ON-state current rating** is usually specified for a half-sine-wave current at a particular frequency. Fig. 15 shows curves of the maximum allowable average ON-state current $I_{TF(avg)}$ for the RCA-2N3873 SCR family as a function of case temperature. Because peak and rms currents may be high for small conduction angles, the curves in Fig. 15 also show maximum allowable average currents as a function of conduction angle. The maximum operating junction temperature for the 2N3873 is 100° C. The rating curves indicate, for a given case temperature, the maximum average ON-state current for which the average temperature of the pellet will not exceed the maximum allowable value. The rating curves may be used for only resistive or inductive loads. When capacitive loads are used, the currents produced by the charge or discharge of the capacitor through the thyristor may be excessively high, and a resist-

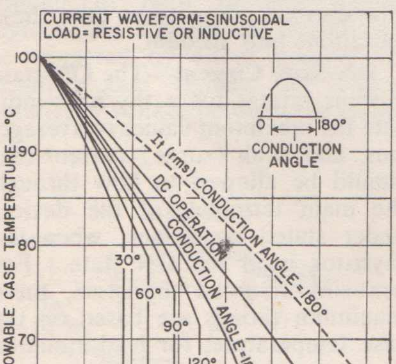


Figure 15. Current rating chart for the RCA-2N3873 SCR.

ance should be used in series with the capacitor to limit the current to the rating of the thyristor.

The ratio of rms to average values for a sinusoidal current waveform through an SCR is 1.57. The maximum average ON-state current rating $I_{TF(avg)}$, therefore, can be readily converted to the maximum rms ON-state current rating $I_{TF(rms)}$. For example, as may be determined from Fig. 15, the maximum average ON-state current for the 2N3873 is 22 amperes for a conduction angle of 180 degrees and a maximum case temperature of 65° C. For these same conditions, the rms current rating may be determined as follows:

$$I_{TF(rms)} = I_{TF(avg)} \times 1.57$$
$$= 22 \text{ amperes} \times 1.57$$
$$= 35 \text{ amperes}$$

The dashed-line curve in Fig. 15 shows the rms current rating for the 2N3873 as a function of case temperature for a conduction angle of 180 degrees.

The ON-state current rating for a triac is given only in rms values because these devices normally conduct alternating current. Fig. 16 shows an rms ON-state current rating

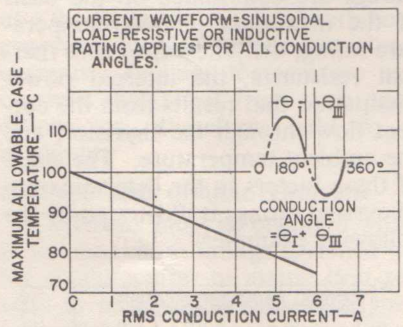


Figure 16. Current rating curve for a typical RCA triac.

curve for a typical triac as a function of case temperature. As with the SCR, the triac curve is derated to zero current when the case temperature rises to the maximum operating junction temperature. Triac current ratings are given for full-wave conduction under resistive or inductive loads. Precautions should be taken to limit the peak current to tolerable levels when capacitive loads are used.

The **surge ON-state current rating** $I_{TF(surge)}$ indicates the maximum peak value of a short-duration current pulse that should be allowed to flow through a thyristor during one ON-state cycle, under stated conditions. This rating is applicable for any rated load condition. During normal operation, the junction temperature of a thyristor may rise to the maximum allowable value; if the surge occurs at this time, the maximum limit is exceeded. For this reason, a thyristor is not rated to block OFF-state voltage immediately following the occurrence of a current surge. Sufficient time must be allowed to permit the junction temperature to return to the normal operating value before gate control is restored to the thyristor. Fig. 17 shows a surge-current rating curve for the 2N3873 SCR. This curve shows peak values of half-sine-wave forward (ON-state) current as a function of overload duration measured in cycles of the 60-Hz current. Fig. 18 shows surge-current rating curves for a typical triac. For triacs, the rating curve shows peak values for a full-sine-wave current as a function of the number of cycles of overload

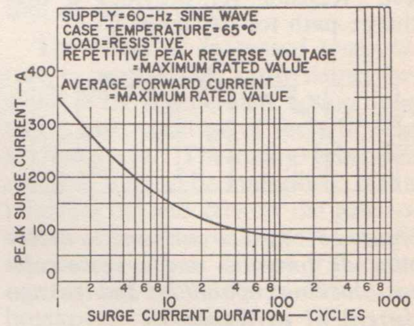


Figure 17. Surge-current rating curve for the RCA-2N3873 SCR.

duration. Multicycle surge curves are the basis for the selection of circuit breakers and fuses that are used to prevent damage to the thyristor in the event of accidental short-circuit of the device. The number of surges permitted over the life of the thyristor should be limited to prevent device degradation.

Critical Rate of Rise of ON-State Current (di/dt)—In a thyristor, the load current is initially concentrated in the small area of the pellet when load current first begins to flow. This small area effectively limits the amount of current that the device can handle and results in a high voltage drop across the pellet in the first micro-second after the thyristor is triggered. If the rate of rise of current is not maintained within the rating of the thyristor, localized hot spots may occur within the pellet and permanent damage to the device may result. The waveshape for testing the di/dt capability of the RCA 2N3873 is shown in Fig. 19. The critical rate of rise of ON-state current is dependent upon the size of the cathode area that begins to conduct initially, and the size of this area is increased

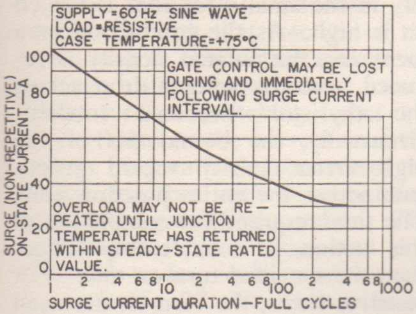


Figure 18. Surge-current rating chart for a typical triac.

for larger values of gate trigger current. For this reason, the di/dt rating is specified for a specific value of gate trigger current.

Holding and Latching Currents—After a thyristor has been switched to the ON-state condition, a certain minimum value of anode current is required to maintain the thyristor in this low-impedance state. If the anode current is reduced below this critical holding-current value, the thyristor cannot maintain regeneration and reverts to the OFF or high-impedance state. Because the holding current (I_H) is sensitive to changes in temperature (increases as temperature decreases), this rating is specified at room temperature with the gate open.

The latching-current rating of a thyristor specifies the value of anode current, slightly higher than the holding current, which is the minimum amount required to sustain thyristor is switched from the OFF state to the ON state and the gate signal is removed. Once the latching current (I_L) is reached, the thyristor remains in the ON, or low-impedance, state until its anode current is decreased below the holding-current value. The latching-current rating is an important consideration when a thyristor is to be used with an inductive load because the inductance limits the rate of rise of the anode current. Precautions should be taken to ensure that, under such conditions, the gate signal is present until the anode current rises to the latching value so that complete turn-on of the thyristor is assured.

Fig. 20 shows a simple test circuit that may be used to determine the holding and latching currents of thyristors. For the holding-current tests, the value of potentiometer R_1 is adjusted to approximately 50

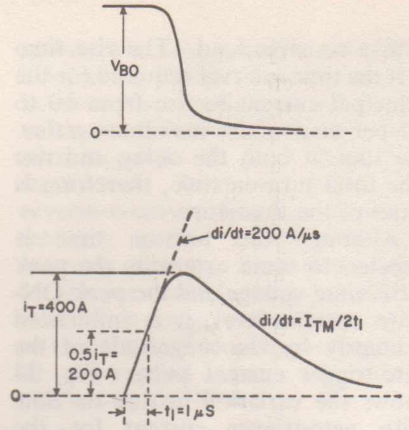


Figure 19. Voltage and current waveforms used to determine di/dt rating of the RCA-2N3873 SCR.

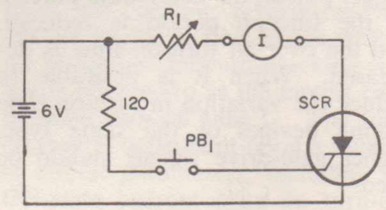


Figure 20. Test circuit used to determine holding and latching currents of thyristors.

ohms, and the spring-loaded push-button switch PB_1 is momentarily depressed to turn on the thyristor. The value of R_1 is then gradually increased to the point at which the thyristor turns off.

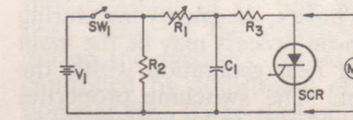
For the latching-current test, the value of potentiometer R_1 is initially adjusted so that the main-terminal current is less than the holding level. The value of R_1 is then decreased, as push-button switch PB_1 is alternately depressed and released, until the thyristor latches on.

Critical Rate of Rise of OFF-State Voltage (dv/dt)—Because of the internal capacitance of a thyristor, the forward-blocking capability of the device is sensitive to the rate at which the forward voltage is applied. A steep rising voltage impressed across the main terminals of a thyristor causes a capacitive charging current to flow through the device. This charging current ($i = Cdv/dt$) is a function of the rate of rise of the OFF-state voltage.

If the rate of rise of the forward voltage exceeds a critical value, the capacitive charging current may become large enough to trigger the thyristor. The steeper the wavefront of applied forward voltage, the lower the value of the thyristor breakover voltage becomes.

The use of the shorted emitter construction in SCR's has resulted in a substantial increase in the dv/dt capability of these devices by providing a shunt path around the gate-to-cathode junction. Typical units can withstand rates of voltage rise up to 200 volts per microsecond under worst-case conditions. The dv/dt capability of a thyristor decreases as the temperature rises and is increased by the addition of an external resistance from gate to reference terminal. The dv/dt rating, therefore, is given for the maximum junction temperature with the gate open, i.e., for worst-case conditions.

Fig. 21 (a) shows a simple test circuit that may be used to determine the dv/dt capability of a thyristor. The curves in Fig. 21 (b) define the critical values for linear and exponential rates of increase in reapplied forward OFF-state voltage for an SCR. The critical value for the exponential rate of rise of forward voltage



V_1 = anode supply voltage (variable)
SW_1 = mercury-wetted relay
R_1 = non-inductive variable resistor
R_2 = discharge resistor
R_3 = current-limiting resistor
M_1 = oscilloscope

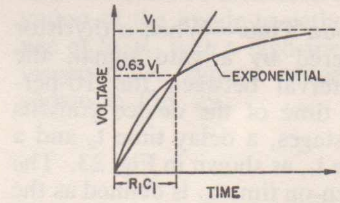


Figure 21. Test circuit and waveforms used to determine dv/dt capability of a thyristor.

is the rating given in the manufacturer's test specifications. This rating is determined from the following equation:

$$\frac{dv}{dt} = \frac{\text{rated value of thyristor voltage (V)}}{\text{RC time constant}} \times 0.632 \quad (7)$$

The dv/dt rating allows a circuit designer to design an RC time-constant network that can be used to limit the rate of rise of a transient voltage below the critical value of the thyristor.

It has been found in many applications that simple circuit additions, such as shown in Fig. 22, can be used to reduce the dv/dt stress on the thyristor. The dv/dt capability is also increased by application of reverse bias to the gate during the rise of OFF-state voltage.

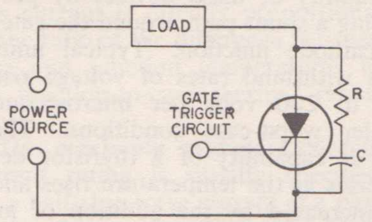


Figure 22. Diagram showing use of RC network to improve the dv/dt capability of an SCR.

SWITCHING CHARACTERISTICS

The ratings of thyristors are based primarily upon the amount of heat generated within the device pellet and the ability of the device package to transfer the internal heat to the external case. For high-frequency applications in which the peak-to-average current ratio is high, or for high-performance applications that require large peak values but narrow current pulses, the energy lost during the turn-on process may be the main cause of heat generation within the thyristor. The switching properties of the device must be known, therefore, to determine power dissipation which may limit the device performance.

Turn-on Time—When a thyristor is triggered by a gate signal, the time interval between the 10-per-turn-on time of the device consists of two stages, a delay time t_d and a rise time t_r, as shown in Fig. 23. The total turn-on time t_{gt} is defined as the time interval between the initiation

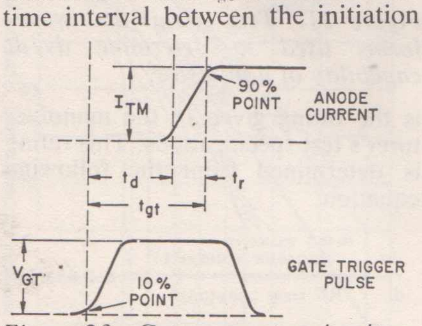


Figure 23. Gate current and voltage turn-on waveforms for a thyristor.

of the gate signal and the time when the resulting current through the thyristor reaches 90 per cent of its maximum value with a resistive load. The **delay time** t_d is defined as the time interval between the 10-per-cent point of the leading edge of the gate-trigger voltage and the 10-per-cent point of the resulting current with a resistive load. The **rise time** t_r is the time interval required for the principal current to rise from 10 to 90 per cent of its maximum value. the sum of both the delay and rise The total turn-on time, therefore, is times of the thyristor.

Although the turn-on time is affected to some extent by the peak OFF-state voltage and the peak ON-state current level, it is influenced primarily by the magnitude of the gate-trigger current pulse. Fig. 24 shows the variation in turn-on time with gate-trigger current for the RCA-2N3873 SCR. When larger currents are available from the gate-trigger pulses, the delay-time portion of the turn-on period is reduced, and the over-all turn-on time is decreased. When it is desirable to reduce the variation in turn-on time among devices of the same type, higher gate-drive signals should be used.

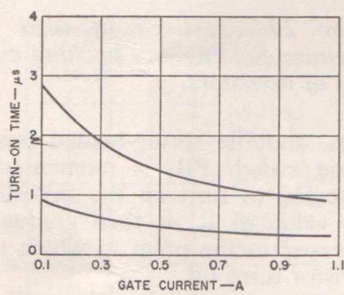


Figure 24. Turn-on time characteristics for the RCA-2N3873 SCR.

Fig. 25 shows a simple test circuit used to determine turn-on times of thyristors. The value of resistor R_1 is chosen so that the rated value of current flows through the thyristor. Turn-on time is specified by the thyristor manufacturer at the rated blocking voltage.

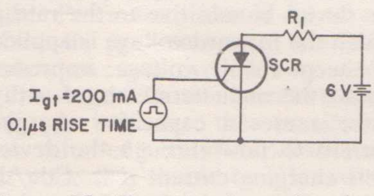


Figure 25. Test circuit used to determine turn-on time of thyristors.

When a thyristor is turned on by a gate-current pulse, current does not start to flow throughout the entire junction instantaneously; instead, the current is confined initially to a small area adjacent to the gate. The voltage drop across the thyristor at this time is large because the current density in the small area that is turned on is high. As the conduction area increases, the current density is reduced, and the voltage drop across the thyristor becomes smaller. Eventually, the boundaries of the high-current-density region propagate across the entire junction area. The time required for completion of this action is considerably longer than the specified turn-on time. For resistive loads, the turn-on time can be defined as the time interval between 10 per cent of the gate voltage and the period required for the applied blocking voltage to decrease to 10 per cent of its original value. For thyristors operated at low blocking voltages, the 10-per-cent value is insignificant from the standpoint of device dissipation. For thyristors operated at blocking voltages in the order of hundreds of volts, however, 10 per cent is sufficiently high in magnitude to represent an appreciable amount of device dissipation. Moreover, the typical turn-on time, as defined for certain gate drives, may be in the order of 2 to 3 microseconds, while the time required for conduction to spread over the entire junction area may be in the order of 20 microseconds. During this spreading time, the dynamic voltage drop is high, and the current density can produce localised hot spots in the pellet area in conduction. In order to guarantee reliable operation and to provide guidance for equipment designers in applications having short conduction periods, the voltage drop across RCA thyristors, at a given instantaneous forward current and at a specified time after turn-on from an OFF-state condition, is given in the published data. The waveshapes for the initial ON-state voltage for the RCA-2N3873 SCR is shown in Fig. 26. This initial voltage, together with the time required for reduction of the dynamic forward voltage drop during the spreading time, is an indication of the current-switching capability of the thyristor.

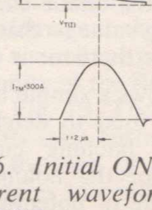


Figure 26. Initial ON-state voltage and current waveforms for the 2N3873 SCR.

When the entire junction area of a thyristor is not in conduction, the current through that fraction of the pellet area in conduction may result in large instantaneous power losses. These turn-on switching losses are proportional to the current and the voltage from cathode to anode of the device, together with the repetition rate of the gate-trigger pulses. The instantaneous power dissipated in a thyristor under such conditions is shown in Fig. 27. The curves shown in this figure indicate that the peak power dissipation occurs in the short interval immediately after the device starts to conduct, usually in the first microsecond. During this time interval, the peak junction temperature may exceed the maximum operating temperature given in the manufacturer's data; in this case, the thyristor should not be required to block voltages immediately after the conduction interval. If the thyristor must block voltages immediately following the conduction interval, the junction-temperature rating must not be exceeded, and sufficient time must elapse to allow the junction temperature to decrease to the operating temperature before blocking voltage is re-applied to the device.

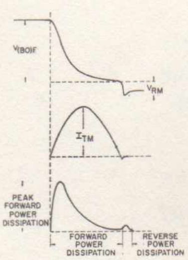


Figure 27. Instantaneous power dissipation in a thyristor during turn-on.

The transient temperature rise may have a major effect on the turn-off time of a thyristor. As a result, when transient effects have to be considered, turn-off time measurements should be made under pulsed conditions.

Turn-Off Time — The turn-off time of an SCR also consists of two stages, a reverse-recovery time and a gate-recovery time, as shown in Fig. 28. When the forward current of an SCR is reduced to zero at the end of a conduction period, application of reverse voltage between the anode and cathode terminals causes reverse current to flow in the SCR until the reverse-blocking junction establishes a depletion region. The time interval between the application of reverse voltage and the time that the reverse current passes its peak value to a steady-state level is called the **reverse-recovery time** t_{rr}. A second recovery period called the **gate-recovery time** t_{gr}, must then elapse for the forward-blocking junction to establish a forward-depletion region so that forward blocking voltage can be re-applied and successfully blocked by the SCR.

The gate-recovery time of an SCR is usually much longer than the reverse-recovery time. The total time from the instant reverse-recovery current begins to flow to the start of the re-applied forward-blocking voltage is referred to as the circuit **commutated turn-off time** t_q. The turn-off time is dependent upon a number of circuit parameters, including the ON-state current prior to turn-off, the rate of change of current during

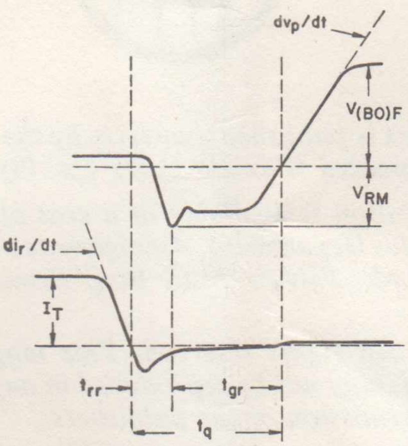
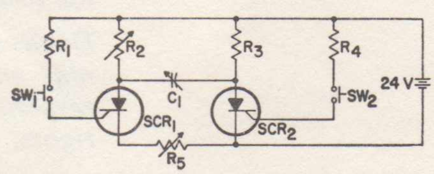


Figure 28. Circuit-commutated turn-off voltage and current waveforms for a thyristor.

the forward-to-reverse transition, the reverse-blocking voltage, the rate of change of the re-applied forward voltage, the gate trigger level, the gate bias, and the junction temperature. The junction temperature and the ON-state current, however, have a more significant effect on turn-off time than any of the other factors. Because the turn-off time of an SCR depends upon a number of circuit parameters, the manufacturer's turn-off time specification is meaningful only if these critical parameters are listed and the test circuit used for the measurement is indicated.

Fig. 29 shows a simple test circuit used to measure turn-off time. The circuit subjects the SCR to current and voltage waveforms similar to those encountered in most typical applications. In the circuit diagram, SCR_1 is the device under test. Initially, both SCR's are in the OFF-state; push-button switch SW_1 is momentarily closed to start the test. This action turns on SCR_1 and load current flows through this SCR and resistor R_2. Capacitor C_1 charges through resistor R_3 to the voltage developed across R_2. If the second push-button switch SW_2 is then closed, SCR_2 is turned on. SCR_1 is then reverse-biased by the voltage across capacitor C_1. The discharge of this capacitor causes a short pulse of reverse current to flow through SCR_1 until this device recovers its reverse-blocking capability. At some time t_1, the anode-to-cathode voltage of SCR_1 passes through zero and starts to build up in a forward direction at a rate dependent upon the time constant of C_1 and R_2. The peak value of the reverse current during the recovery period can be controlled by adjustment of potentiometer R_5. If the turn-off time of SCR_1 is less than the time t_1, the device will turn off. The turn-off interval t_1 can be measured by observation of the anode-to-cathode voltage of SCR_1 with a high-speed oscilloscope. A typical waveform is shown in Fig. 29.



$R_1 = R_4 = 100$ ohms
R_2 = variable resistor, 0.7 to 50 ohms
$R_3 = 5000$ ohms
R_5 = variable resistor, 0.1 to 1 ohm
C_1 = variable capacitor, 0.1 to 1 μF, 150 V
SCR_1 = SCR under test
SCR_2 = RCA-40378

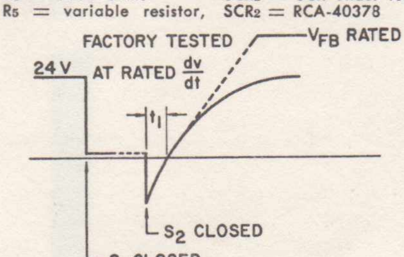


Figure 29. Test circuit and voltage waveforms used to determine turn-off times of thyristors.

(TO BE CONTINUED)

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